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BINARY PSEUDO RANDOM PHASE CORRELATION
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Authors: M Dean & I Barrow

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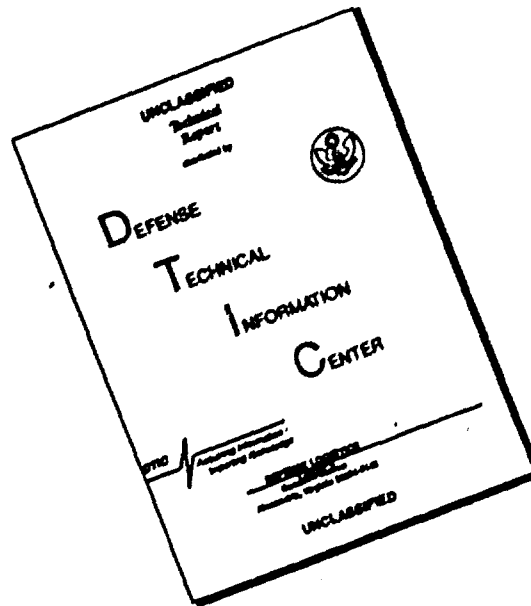
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AUTHORS: M DEAN AND IAN BARROW

DATE : JANUARY 1989

SUMMARY

Contd from Doc

Techniques for achieving the correlation of binary pseudo random phase coded signals in doppler radar sensors are discussed. A new design based on current technology is developed which potentially offers both a low cost and highly integrable solution. Prototype hardware is constructed and tested and its performance compared with other designs. Excellent performance is achievable without the penalties of size and weight which the other designs carry.

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BINARY PSEUDO RANDOM PHASE CODE CORRELATION IN RADAR SYSTEMS -
A NEW APPROACH

1 INTRODUCTION

The growing requirement for low probability of intercept characteristics in radar sensors is producing increasing interest in continuous wave (cw) radars. Pure cw radars have no inherent ranging capability, and it is therefore customary to modulate the cw transmission in some way if ranging is required.

An increasingly popular modulation scheme is Binary Pseudo Random Phase Coding (BPRPC), whereby the phase of the transmitted signal is switched between 0° and 180° under the control of a binary pseudo random sequence of length n, variously known as a pseudo random code, maximal sequence or Galois Code.

A detailed review of this form of modulation is given in reference 1. Also included in reference 1 is a treatment of some of the practical difficulties of implementing such a modulation, particularly as regards demodulating the received target returns.

This can be accomplished by feeding the returns into a transversal filter, as shown in figure 1, with weighting coefficients of +1 and -1 corresponding to the phase reversals of the transmitted code. The device is an "all range" correlator, in that all range delays within the code repeat interval are processed by the one device, and may be realised as a digital filter or using analogue surface acoustic wave technology. However, problems of dynamic range in the former and lack of programability in the latter mean that for many applications a different technique is required.

One such alternative technique, often known as an active correlator, is illustrated in figure 2. The correlation of the received binary phase coded signals is accomplished by a simple analogue balanced mixer. In essence this is a switched inverter controlled by a delayed replica of the transmitted signal. In practice the correlation is normally accomplished after mixing down to video frequencies. The output of the circuit is a baseband version of the received signals, which can be fed directly to a doppler filter bank, or via a clutter pre-filter.

This circuit is capable of very high dynamic range operation, and is independent of code modulation. The circuit is also intrinsically capable of operating at very high code rates. It has the disadvantage, however, that a separate circuit is required for each implemented range gate. This problem is exacerbated by the presence of transformers and baluns, which cannot be simply produced in integrated form. A programme of work was accordingly initiated to find implementations of the active correlator circuit of figure 2, which can be implemented in a compact, low cost, integrated form. The results of this study form the subject of section 3 this paper.

2 THE DIODE/TRANSFORMER MIXER AS AN ACTIVE CORRELATOR

2.1 Principle of Operation

The internal architecture of a diode/transformer mixer suitable for correlation of binary pseudo random coded signals is shown in figure 3. When used as a binary phase code correlator the voltage on the IF terminal is switched between positive and negative by the pseudo random sequence. If the IF terminal is positive with respect to earth the LO and RF transformers are connected via diodes 1 and 3. If the IF terminal is negative the transformers are connected via diodes 2 and 4. We see that the sense of the connections between the two transformers is reversed in sympathy with the code. Hence any input signal from the LO port will be transformed through to the RF port with either zero or 180° phase shift, depending upon the polarity of the code applied at the IF port, thus performing the required correlation action.

Some commercial analog mixers allow access to both centre taps on the diode ring transformers. The above effect may hence be achieved by applying binary code (0,1) and inverse code signals to the two respective centre taps. Using this method it is not necessary to generate negative code voltages, hence the whole system can be more easily implemented with standard digital technology.

2.2 The Diode/Transformer Mixer As An Active Correlator - Measured Performance

In order to assess the performance of a number of commercially available low cost diode transformer mixers as active correlators a set of experimental evaluations was conducted.

The apparatus used to conduct these evaluations is shown in figure 4. Each mixer was fed by a pseudo random code sequence and a frequency shifted version of the same sequence to represent a doppler target return. The frequency of the false doppler was varied between 0 and 100KHz and the output fed to a spectrum analyser slaved to the false doppler generator.

The devices tested are listed below.

Mini Circuits	SRA8	Diode/Transformer Mixer
"	"	SRA6
"	"	TFM-3
"	"	RMS-1

2.2.1 Mini Circuits SRA8

This diode mixer is ideally suited to correlation. Its LO and RF ports operate from 0.5 kHz to 10 MHz, and the IF port from DC to 10 MHz. Since the mixer allows access to both IF ports it is easily driven by the normal and inverted outputs of a TTL gate, ie code and inverse code. The input signal is applied at the RF port and the correlated output taken from the LO port. This is somewhat unconventional for an analogue mixer, but works well in practice.

A diagram of the circuit configuration used is given in figure 5.

The low and full bandwidth frequency responses for the SRA8 are given in curves (b) of figures 6 and 7 respectively.

We see from the low frequency response that the correlated output has a 3 dB cut-off at approximately 0.5 kHz, which is in accordance with the mixer specification. The full bandwidth response shows a flat response out to at least 100 kHz.

2.2.2 Mini Circuits SRA6

This is a very similar device to the SRA8. Its LO and RF port frequency responses are specified as 3 kHz to 10 MHz, and the IF port from DC to 10 MHz. The drive circuit is identical to that of the SRA8 (figure 5). The low and full bandwidth frequency responses, figures 6 and 7 curves (c), show that the lower 3 dB cut off is approximately 1.5 kHz, flattening with increasing frequency.

2.2.3 Mini Circuits TFM-3

The TFM-3 is a much smaller device (0.5" x 0.25" x 0.23") than either of the SRA models, but has the disadvantage that the single IF port requires positive and negative voltages, hence a modified drive circuit is required. In fact the mixer is used in its conventional form with the output being taken from the IF port and bi-polar code applied to the LO port.

The frequency response of the TFM-3 is specified with LO and RF ports from 40 kHz to 400 MHz and IF port from DC to 400 MHz. The low and full bandwidth frequency responses of the TFM-3 used as a correlator are shown in figures 6 and 7, curves (d). A very flat response is obtained over the full band.

2.2.4 Mini Circuits RMS-1

The RMS-1 is a surface mount mixer, which gives it a considerable compactness advantage (0.25" x 0.31" x 0.2") over the previous analog mixers. Its frequency specification is 0.5 to 500 MHz for LO and RF, and DC-500 MHz for IF. When driven using the same configuration as for the TFM-3, that is with the IF (DC-500 MHz) response taken as output, the responses obtained are as shown in figures 6 and 7 curves (e). It can be seen that in this configuration the amplitude versus frequency response shows an approximately constant insertion loss of 6-7 dBs.

Other mixers, such as the SRA8, showed an insertion loss of only 1-1½ dBs. The reason for the difference can be explained by the poor low frequency response of the LO and RF ports. If we consider the spectrum of a typical pseudo random binary phase code, having a clock frequency of f_c and length n , we find the spectrum consists of lines equally spaced by f_c/n with a "sinc" distribution in amplitude and a first null at f_c . The signal input has a similar spectrum, but is displaced by an amount equal to the doppler frequency. If the mixer has a poor low frequency response the energy in the lower frequency spectral lines is inefficiently coupled to the output, leading to the observed increase in insertion loss.

When driven using the same configuration as for the SRA6 and SRA8 models (figure 5) the frequency response obtained was as shown in figure 7 curve (f). The high pass characteristic exhibited is due to the high pass response of the LO port of the mixer.

2.3 The Diode Transformer Mixer As An Active Correlator - Conclusions

The majority of the mixers tested work extremely efficiently as active correlators. Their main disadvantages are their size and weight and the extreme difficulty of incorporating them into an integrated circuit. For a radar possessing many range gates the size and weight penalty would be considerable. The latest surface mount designs are a considerable improvement in this respect, albeit at the cost of some loss in efficiency due to their limited low frequency response. The integration problems remain, however.

It was concluded that a desirable alternative solution would be a scheme, using solid state integrated circuit techniques, which would provide the same function as the analogue mixer, but not have its inherent size and weight penalties. This, if achievable, would also allow for integration of other receiver components. As an example, the correlator, clutter filter and amplifiers for a single range gate could all be incorporated on a single silicon chip. We will now consider possible schemes for active correlation using integrated circuit techniques.

3 SOLID STATE ACTIVE CORRELATORS

3.1 Differential Amplifier and Analogue Switch

One possible simple scheme for realising an integrated circuit active correlator is illustrated in figure 8. It comprises a differential amplifier followed by a double pole switch.

Consider the operation of the circuit as a binary phase correlator. An input signal applied to the differential amplifier produces outputs which have 180° phase difference between them. By sampling these two signals in synchronisation with a delayed version of the transmitted code the modulating code phase shifts can be removed from the received signal, thus producing the desired baseband doppler output.

The speed limitations of this circuit are a function of the bandwidth of the differential amplifier and the switching speed of the analogue switch. Common differential amplifiers, such as the LM733, offer bandwidths in the region of 90 MHz. Low cost analogue switches have significant switching times between their on and off states, typically 50 ns for the Radio Spares HI200. In practice this means that, for the rise and fall times of the switch, the output is a combination of the two input signals. This effect results in an overall loss of signal due to the switching period. If the switching period is a significant proportion of the clock cycle a large percentage of the signal will be lost, leading to a poor correlation efficiency.

Faster switches have recently become available, with switching speeds of less than 5 ns, however these are relatively expensive, costing in the region of £30 each. None were available for testing at the time this report was written, but presumably they would operate at much higher clock frequencies than the lower cost switches.

A hardware version of the circuit of figure 8 was constructed and tested using the LM773 differential amplifier and HI200 analogue switch. At low clock frequencies, of the order of 10's of kHz, the circuit operated with a small conversion loss. As the clock frequency was increased, into the high 100's of kHz region, the conversion loss became significant, with an effective usable upper limit of 1MHz.

3.2 Silicon Balanced Modulator/Demodulator

This circuit provides an alternative realisation of an all solid state active correlator. It consists of an upper quad differential amplifier driven by a standard differential amplifier with dual current sources (figure 9). The output collectors are cross coupled so that full wave balanced multiplication of the two input voltages occurs. The device therefore lends itself to balanced modulation or active correlation. A silicon integrated circuit implementation of this circuit is readily available, at low cost, in the form of the MOTOROLA type MC1496. These are currently sold at 50 pence each.

The upper quad differential amplifier has limited linear signal handling capabilities, since the transistors have no emitter resistance. Hence the maximum input voltage for linear operation is 26 mV. This factor makes the upper quad more suited as the code input, which is essentially binary. The lower differential amplifier has provision for an external bias resistance R_E , hence the maximum linear input signal may be extended to $V = I_5 R_E$, where I_5 is the bias current. This extended linear range makes this input well suited to the signal input for the correlator. A test circuit was accordingly constructed using the configuration of figure 10. Experimental use of this circuit as a correlator gave good conversion efficiency and good suppression of unwanted code related outputs over a wide range of input signal amplitudes.

From the results obtained using the two alternative types of device, ie the differential amplifier plus analogue switch and the silicon balanced modulator, the latter offered better performance at lower cost. It was concluded, therefore, that this approach appeared most suited to the current application of producing an all solid state correlator.

4 OPTIMISATION OF A CORRELATOR USING A SILICON BALANCED MODULATOR/DEMODULATOR (MC1496)

General data on the MC1496 device is available from reference 2. Whilst testing the circuit of figure 10 as a coder/correlator several adverse characteristics were observed. These comprised clock and code breakthrough and loss of output signal with removal of code. Each of these three points will now be discussed in more detail.

4.1 Clock Breakthrough

Operating the MC1496 as a correlator, it was noticed that the rise and fall times on the code switching edges were significant. The recommended operating current of 1 mA was found, in practice, to be too low to achieve fast switching. Through experimentation a current of 3.2 mA was found to be more acceptable. R_E in figure 10 was accordingly reduced to 3 K Ω .

The load resistors R_L were also reduced to 3K Ω to improve the switching speed. This also had the effect of reducing the gain. Optimum bandwidth should be achieved with a gain of approximately one. Therefore the resistance R_e was also made equal to 3K Ω , as

$$\text{gain } G = \frac{R_L}{R_E + 2r_e} \quad \text{where } r_e = \frac{26 \text{ mV}}{I_5 \text{ mA}} = 8.125\Omega$$

Using the new values of R_L , R_E and R_e , an improved performance in terms of clock breakthrough was achieved. The clock breakthrough was now approximately 20% of its original value.

4.2 Code Breakthrough

It was observed that applying a coded signal to the signal input terminal had an adverse effect with respect to the level of code lines appearing in the output spectrum. The cause of this appeared to be the upsetting of the dc conditions on the two transistors in the lower differential amplifier of figure 9.

Both inputs are connected to ground through 51Ω resistors. These resistors form potential dividers with the $10K\Omega$ resistors and the $50K\Omega$ potentiometer, hence each input has a negative offset, the level of which may be varied by the potentiometer. By dc coupling on to the signal input, one 51Ω resistor is effectively short circuited and puts the input to dc ground. Since the other input, pin 4, will always have a negative offset this means there is a dc differential voltage on to the differential amplifier. Referring to the circuit of figure 11, the current amplifier offers a low output impedance to both dc and ac signals alike, typically a few ohms. By including a 47Ω series resistor, the impedance looking back into the amplifier can be made equivalent to 50Ω . This means that the dc balance on inputs pins 1 and 4 will be maintained. The current amplifier also acts as an impedance buffer. It was further found that using two resistors of equal value, as opposed to the split resistance of the potentiometer, provided adequate dc balance. Significant rejection of code breakthrough was achieved without any further adjustment. Removing this potentiometer assists in the integration of the final design.

4.3 Differential Code Input

With the carrier (code) input ac coupled as in figure 10, a differential input signal is only present on the upper quad of the MC1496 when the code is operational. A simple modification allows the circuit to be used in both coded and uncoded pure cw radar modes without any extra switching.

This modification is shown in figure 12, and replaces the components connected to pins 7 and 8 of the MC1496 in figure 10. If the radar is working in an uncoded pure cw mode it is only necessary to arrange that the code input is replaced by a fixed dc level. Under these conditions the correlator works automatically as a simple amplifier, and passes any signals present for doppler spectral analysis.

5 CHARACTERISATION OF THE MC1496 CORRELATOR

Having optimised the circuit with respect to code and clock breakthrough and cw operation, it remains to measure the dynamic range, frequency response and usable clock bandwidth of the circuit. The revised circuit is shown in figure 13.

5.1 Dynamic Range

This was measured using the configuration of Figure 14. For the purposes of this paper we will define the dynamic range as the difference between the noise floor and the 1 dB compression point. Initially the noise floor measurement was limited by the noise floor of the spectrum analyser used at approximately -90 dBm. To overcome this problem a low noise amplifier was inserted between the output of the correlator and the input to the spectrum analyser (figure 14) to establish a lower overall noise floor. Using this method the output of the correlator was measured with linearity down to -100 dBm before reaching the noise floor.

To measure the 1 dB compression point of the correlator, a power amplifier was inserted directly between the code modulator and the input attenuator (figure 15). This power amplifier enabled the input power to the correlator to be increased such that the 1 dB compression point could be measured. We see from the results in Table 1 and figure 16 that a 1 dB compression point of $+14.3$ dBm was obtained.

This gives an overall dynamic range of greater than 110 dB which is more than adequate for most radar applications.

5.2 Input Impedance

The use of a input buffer amplifier, such as the LH0002, is desirable to increase the signal input impedance of the correlator. This is because many correlators will be connected in parallel, thereby reducing the overall input impedance and creating drive problems for the preceding amplifier stage.

5.3 Frequency Response

In order that the correlator be suitable for radar application it must have a doppler frequency response at least equal to that of the clutter filters which immediately follow it. The doppler frequency response of the experimental correlator was measured using the circuit of figure 4 and is shown in figures 6 and 7 as curves (a). The response is substantially flat from 0.5 kHz to at least 100 kHz, which is more than adequate for the majority of radar applications.

5.4 Clock Bandwidth

The achievable clock bandwidth is an important parameter of the correlation circuit, as it determines the final radar range resolution achievable. In order to measure the performance of the solid state correlator as a function of clock frequency the clock frequency input to the code generator of figure 4 was varied. No significant deterioration in the performance of the correlator was noted as the clock was increased up to the 14 MHz limit imposed by the code generator. This indicates that the correlator will function adequately in the majority of radar applications.

6 CONCLUSIONS

A silicon balanced modulator has been configured as a binary pseudo random phase code correlator, and has been shown to give performance characteristics adequate for many radar applications. This offers an extremely compact and low cost solution to the correlator problem of phase coded radars. More importantly, it is a semiconductor based design which makes total integration of receiver signal processing components on to a single substrate feasible.

REFERENCES

- 1 DEAN, M. "Binary Phase Coded Radar. A review with suggestions for Future Research". RSRE Memo No 3906, May 1986
- 2 Motorola Semiconductors 1981/82 Linear IC's pp6-76 to 6-85.

INPUT POWER dBm (PHASE CODED)	OUTPUT POWER dBm (CORRELATED)
0	-1.9
5	3.1
10	8.0
11	9.0
12	9.8
13	10.8
14	11.7
15	12.6
16	13.4
17	14.1
18	14.5
19	14.7
20	14.9

TABLE 1 - MEASUREMENT OF 1dB
COMPRESSION POINT

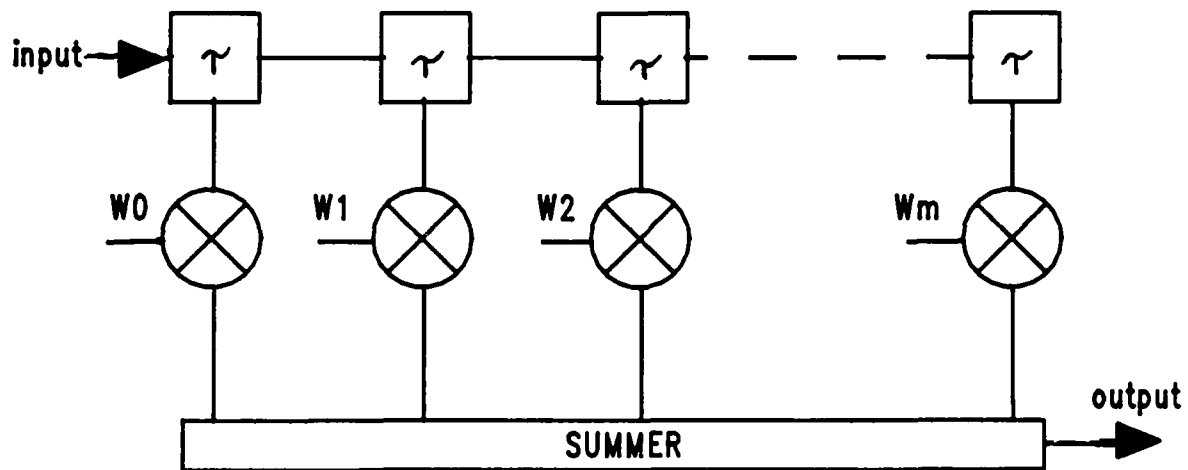


FIG.1 - TRANSVERSAL FILTER

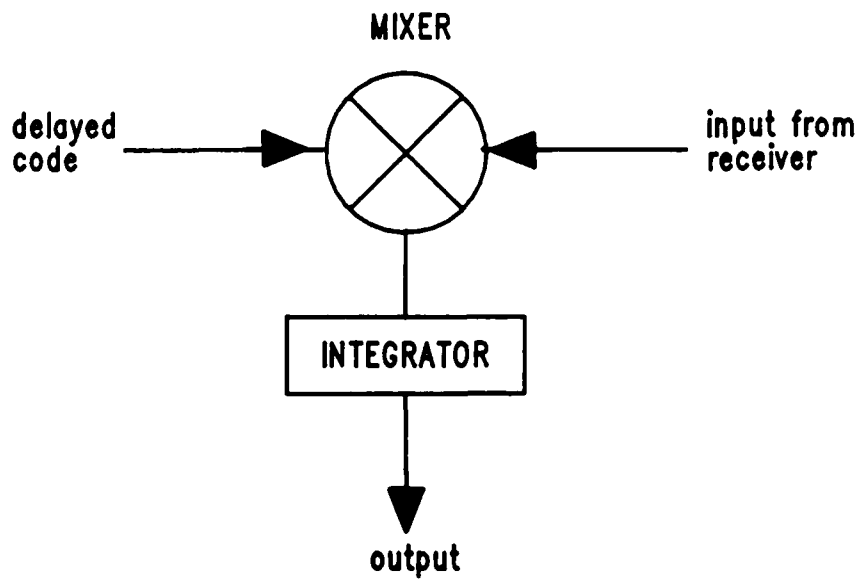


FIG.2 - ACTIVE CORRELATOR

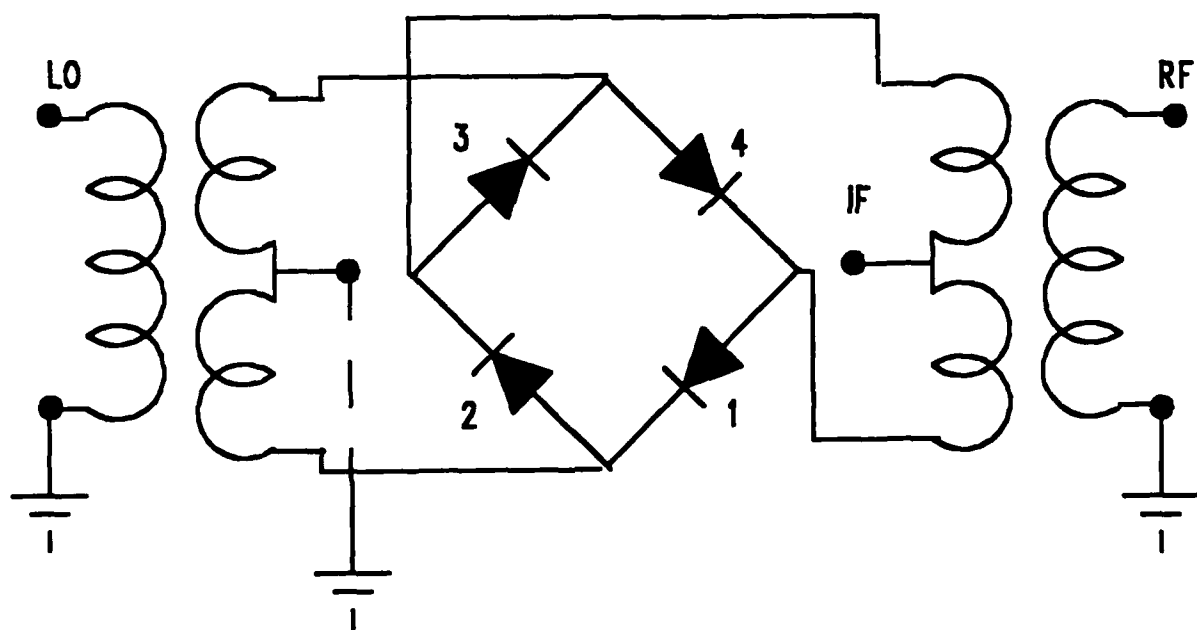


FIG.3 - TRANSFORMER-DIODE TYPE MIXER

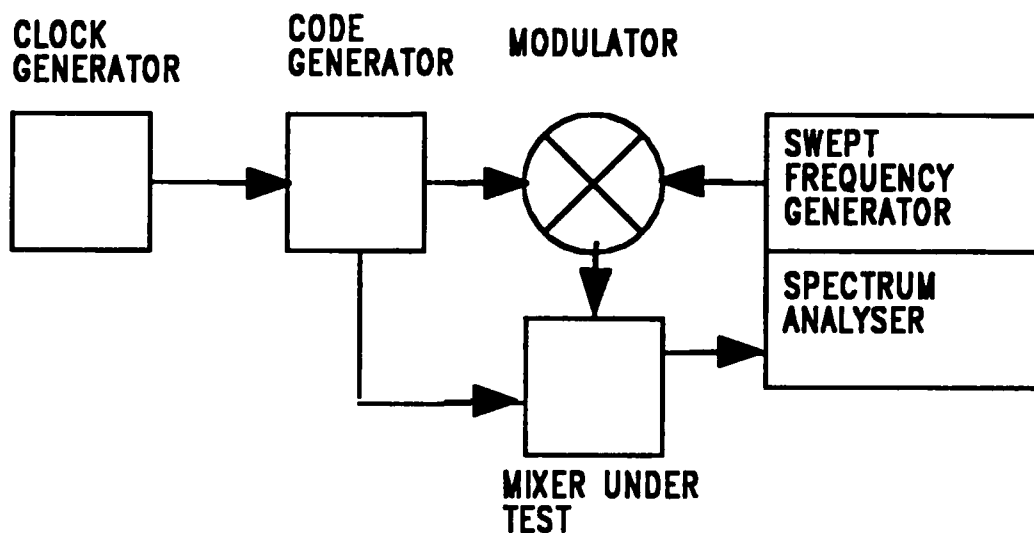


FIG.4 - CONFIGURATION FOR MEASURING
FREQUENCY RESPONSE

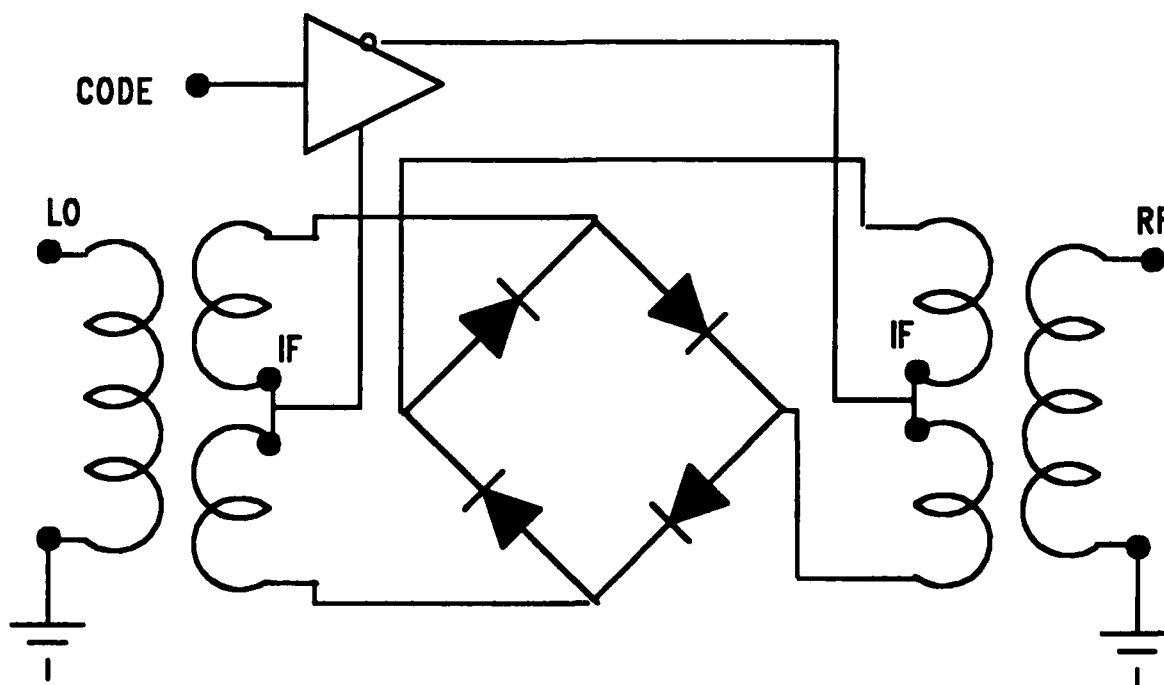


FIG.5 - METHOD FOR DRIVING MIXER WITH
ACCESS TO BOTH IF PORTS

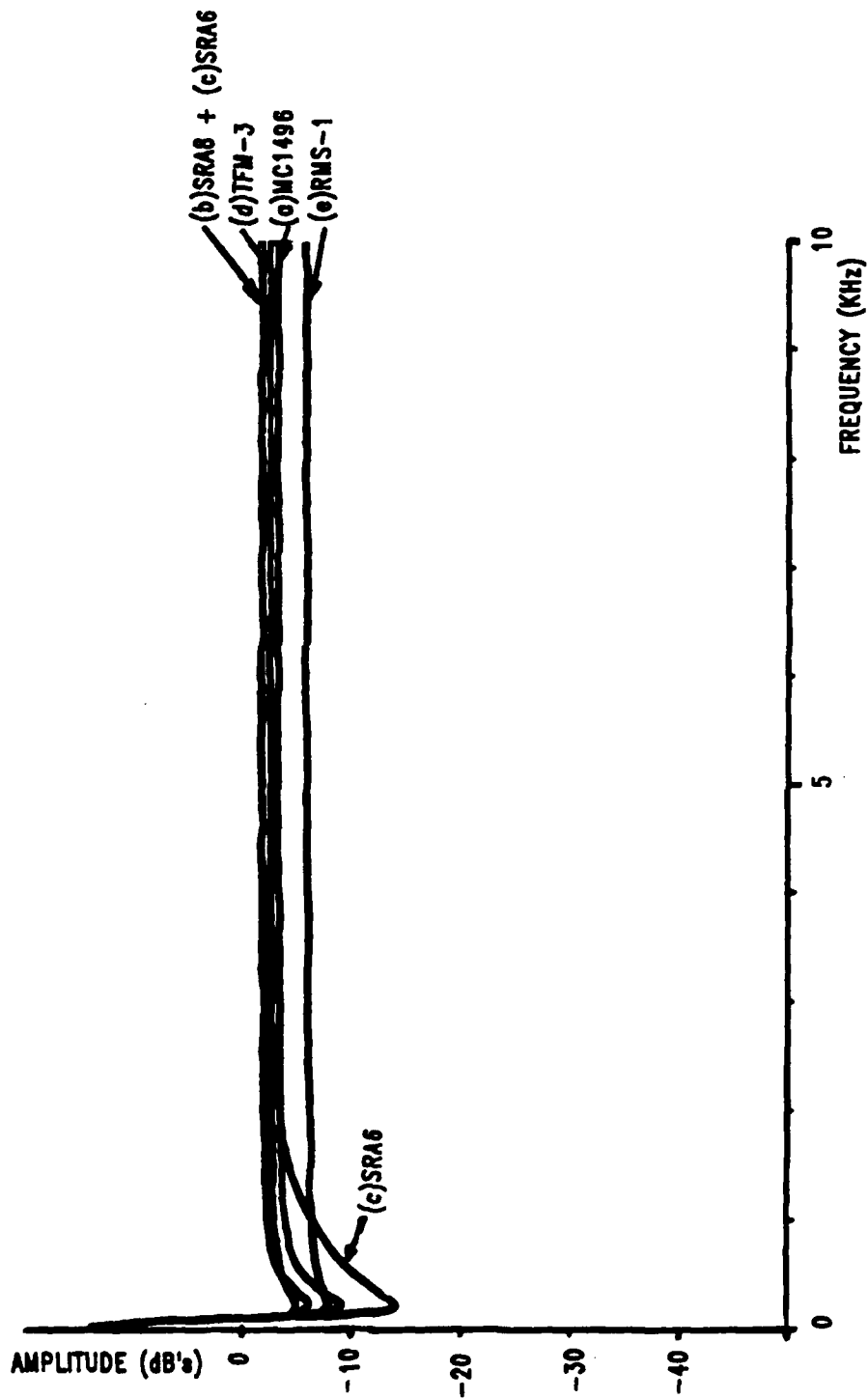


Figure 6 Low Frequency Response of Test Correlators

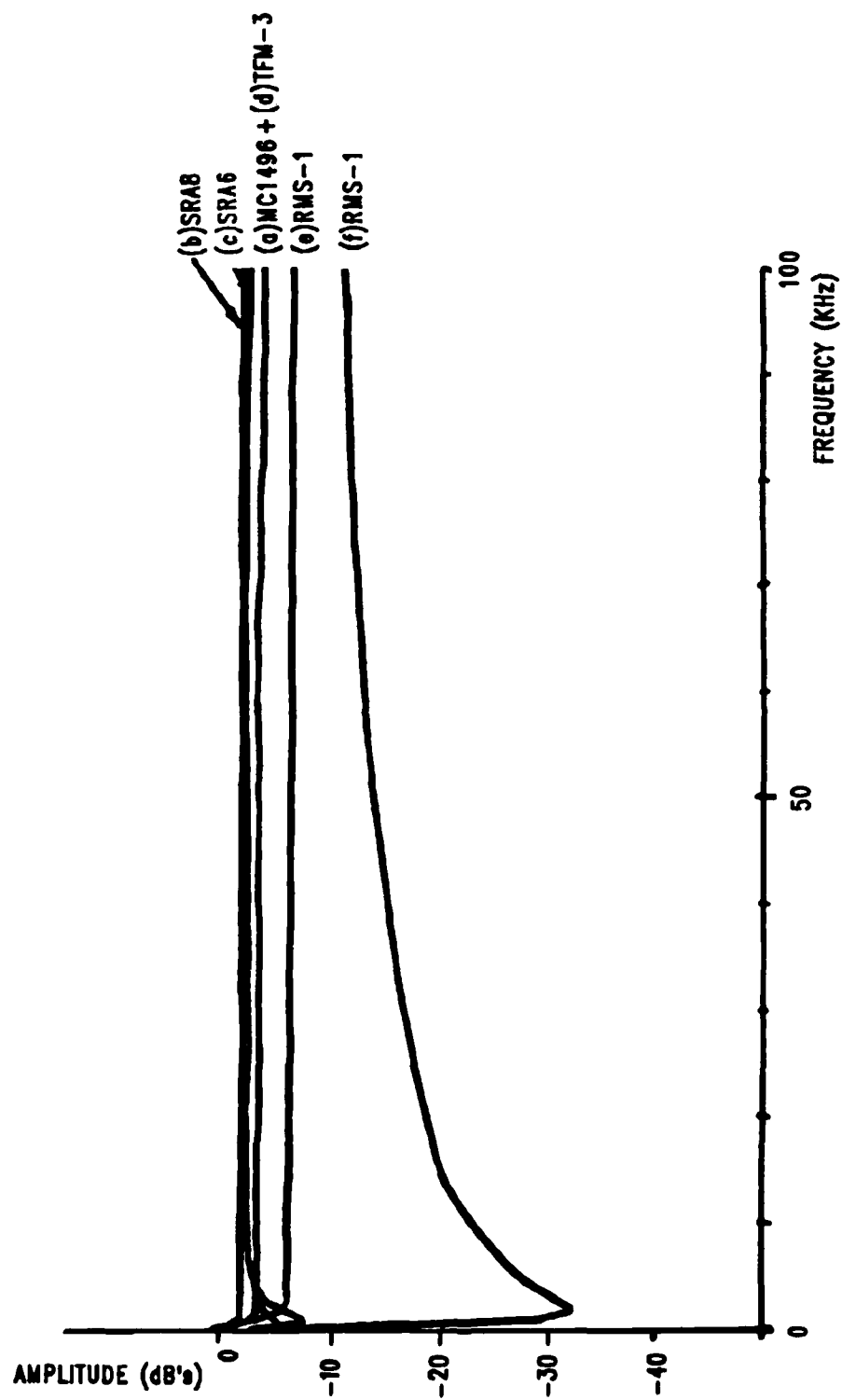


Figure 7 Full Frequency Response of Test Correlators

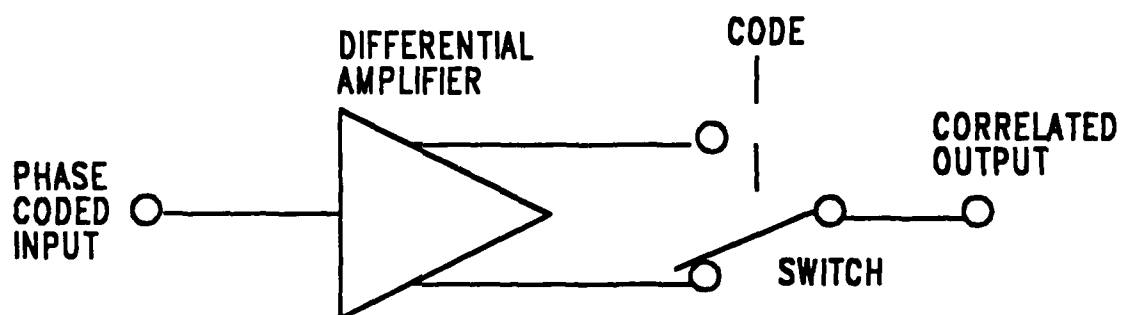


FIG.8 - DIFFERENTIAL AMPLIFIER + SWITCH
TYPE CORRELATOR

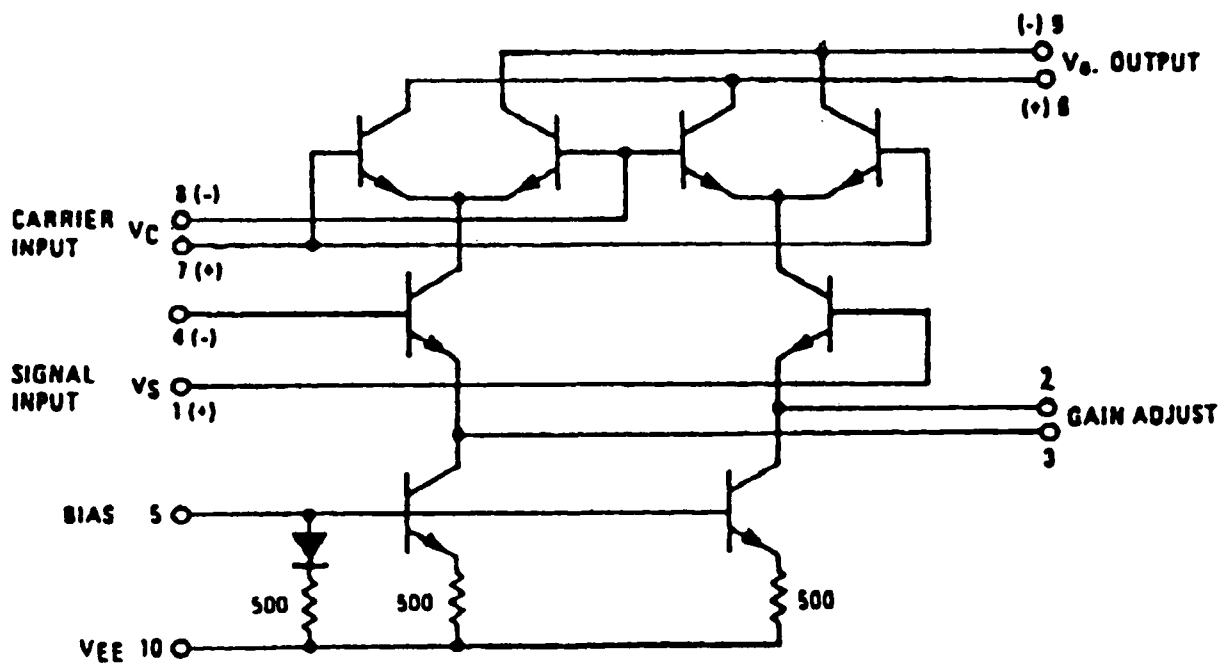


Figure 9 - MC1496 Internal Architecture

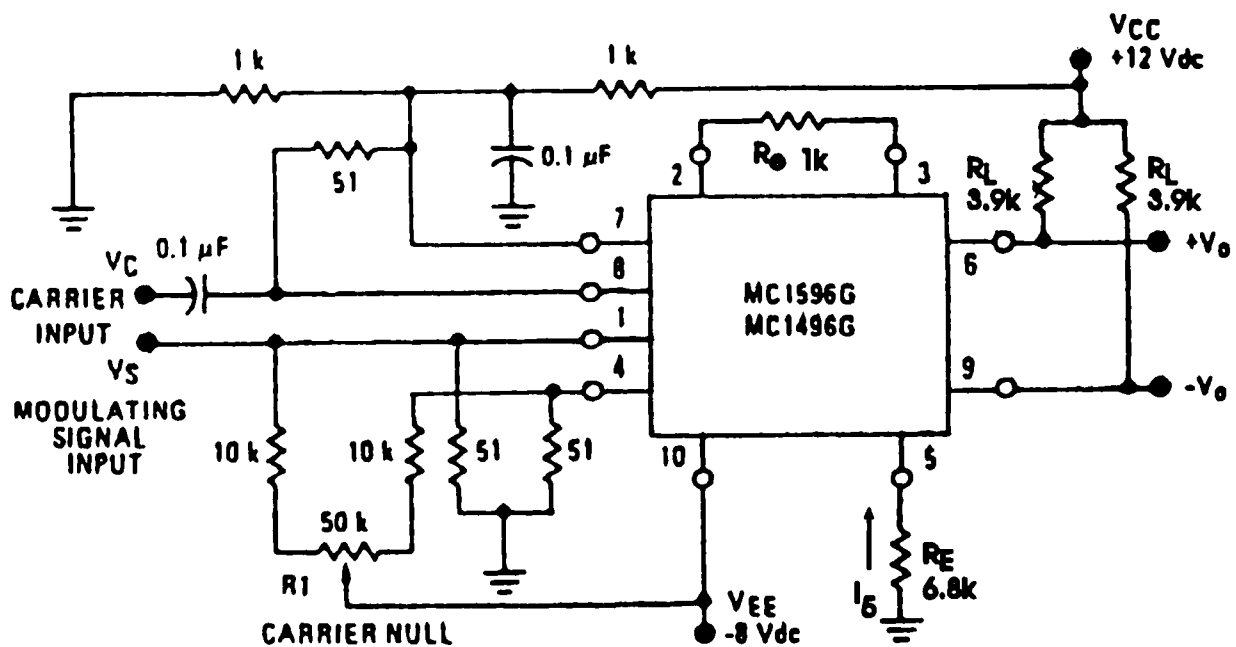
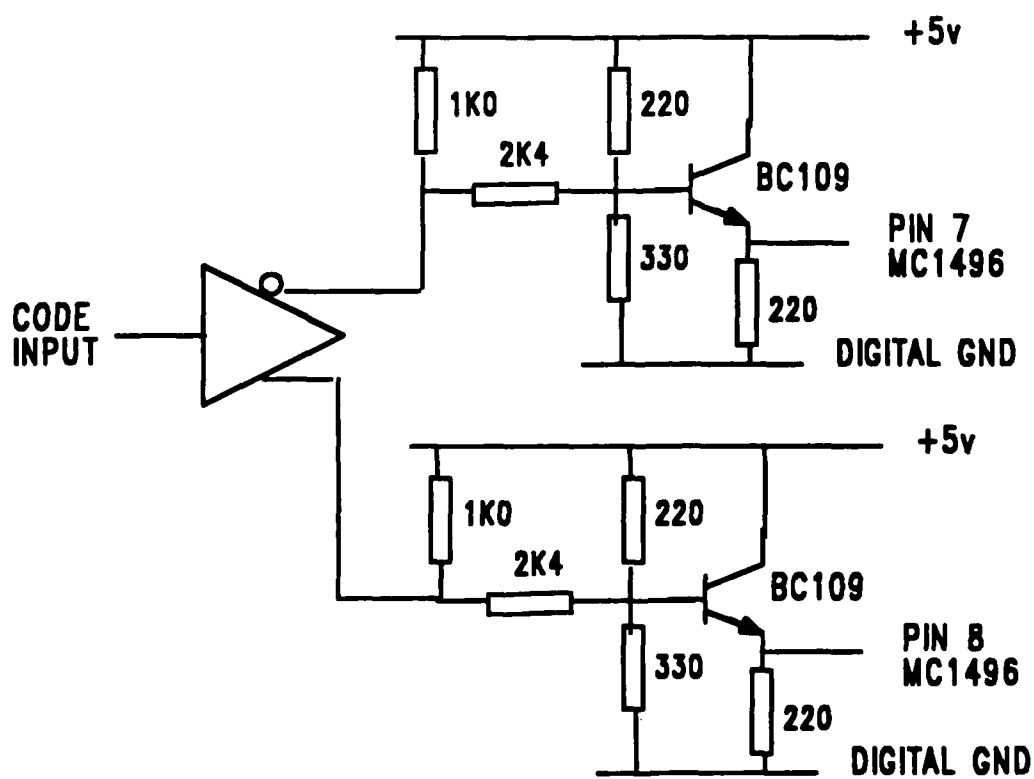
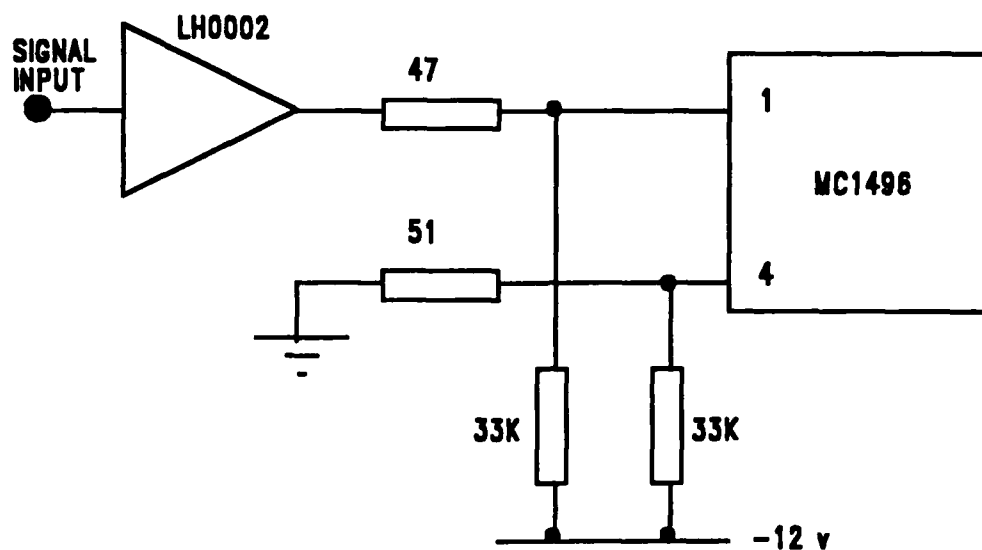


Figure 10 - MC1496 Configured as an Active Correlator



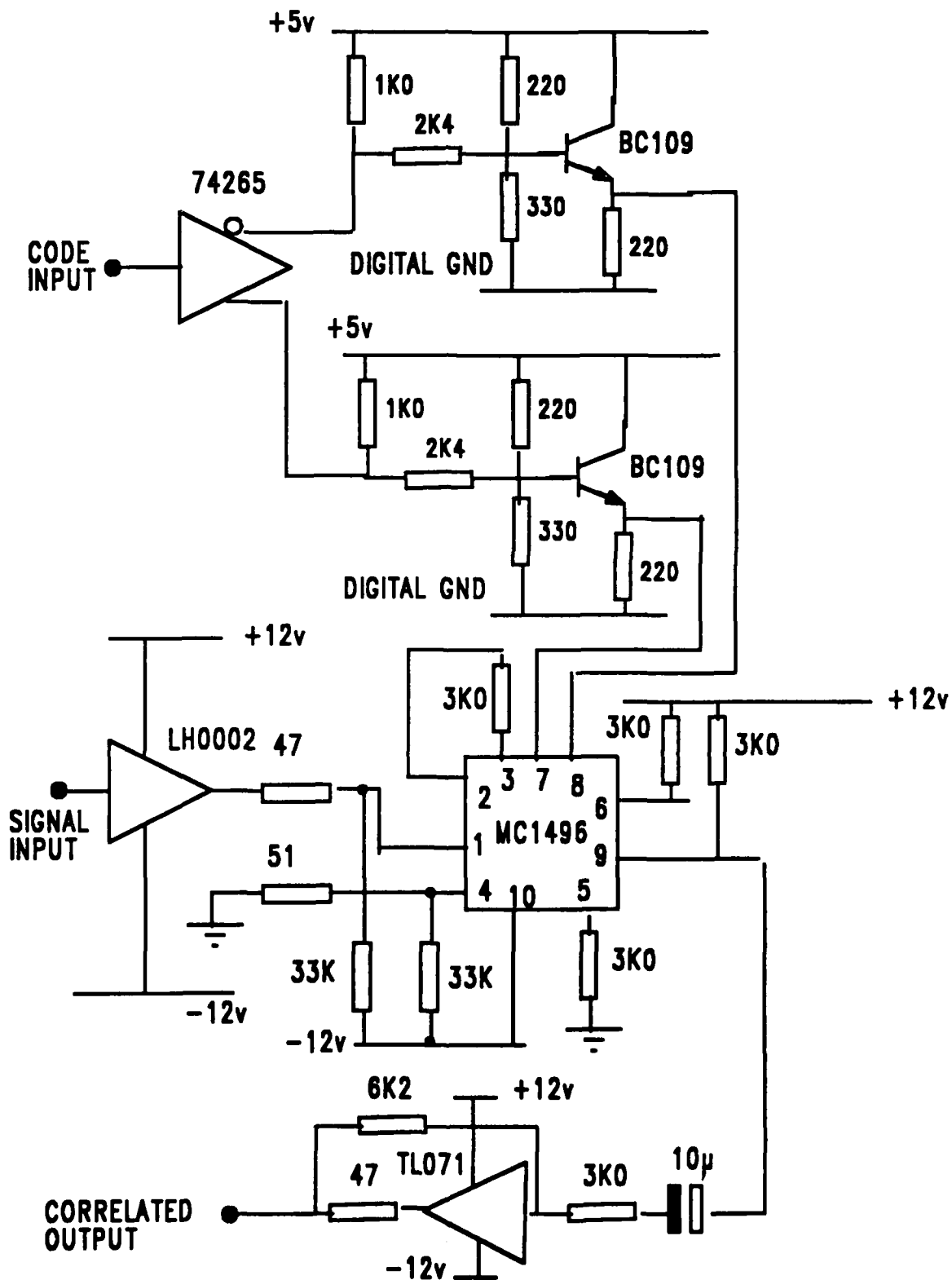


FIG.13 - FINAL MC1496 CORRELATOR DESIGN

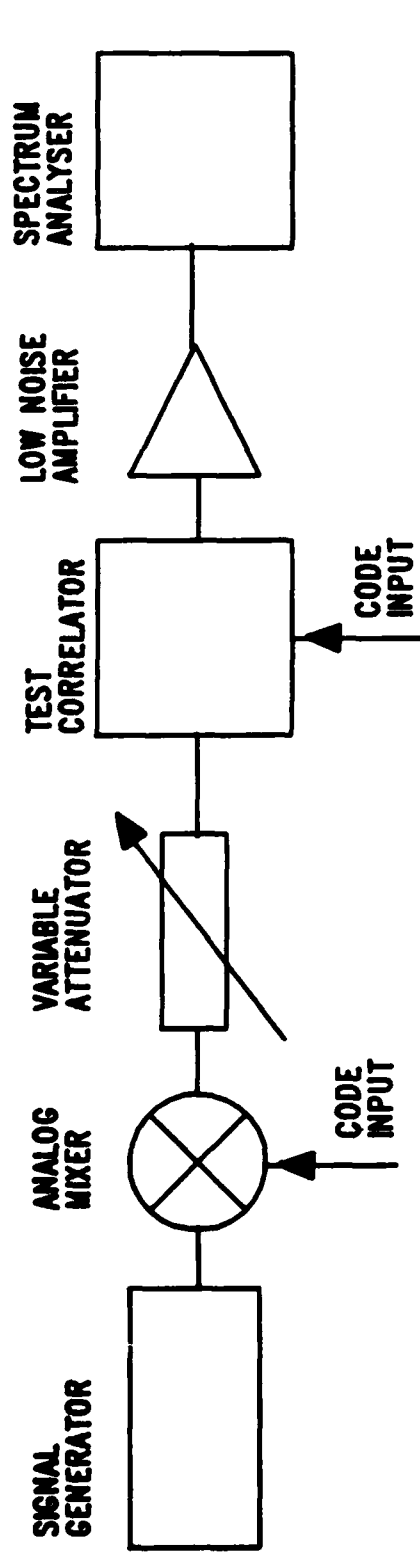


FIG.14 - CONFIGURATION FOR MEASURING NOISE FLOOR OF CORRELATORS

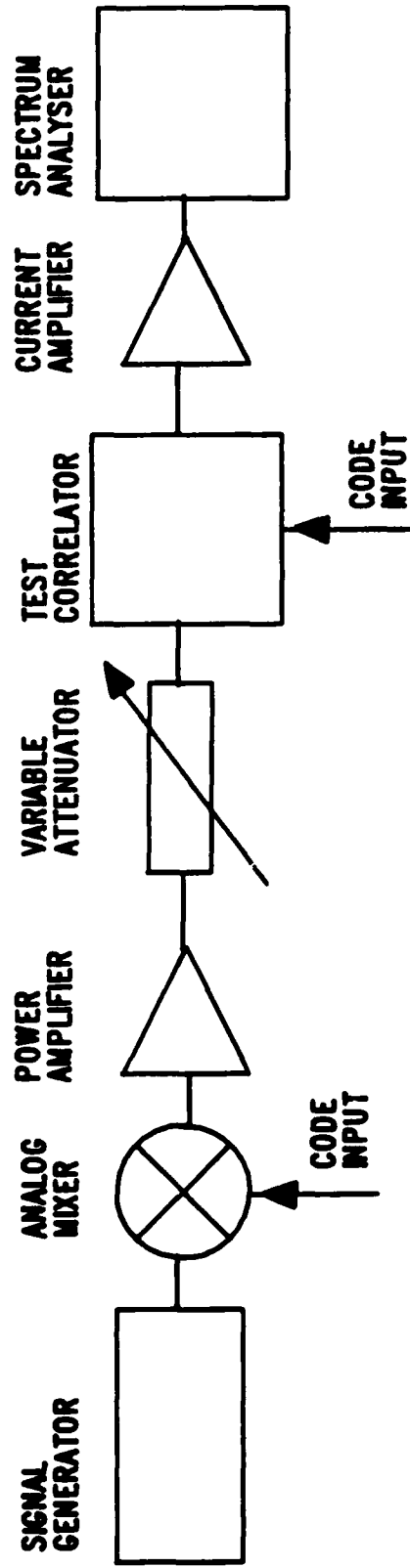
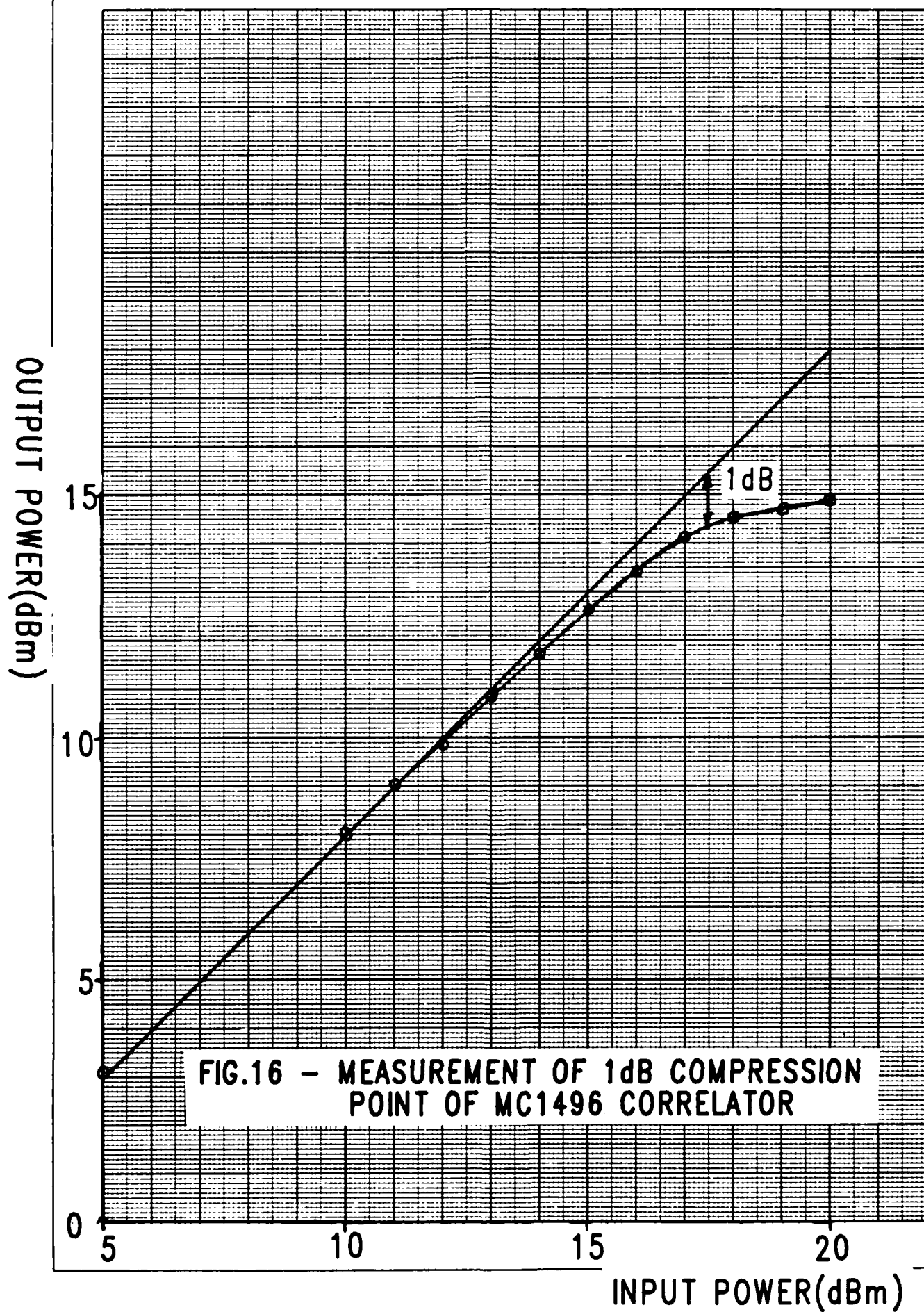


FIG.15 - CONFIGURATION FOR MEASURING 1dB COMPRESSION POINT OF CORRELATORS



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7b. Presented at (for conference papers) Title, place and date of conference				
8. Author 1 Surname, initials DEAN M	9(a) Author 2 BARROW 1	9(b) Authors 3,4...	10. Date 1989.01	pp. ref. 20
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Abstract Techniques for achieving the correlation of binary pseudo random phase coded signals in doppler radar sensors are discussed. A new design based on current technology is developed which potentially offers both a low cost and highly integrable solution. Prototype hardware is constructed and tested and its performance compared with other designs. Excellent performance is achievable without the penalties of size and weight which the other designs carry.				

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